

## CMOS REFERENCE VOLTAGE CIRCUIT

[0001]

### FIELD OF THE INVENTION

5           This invention relates to a reference voltage circuit. More particularly, it relates to a CMOS reference voltage circuit which is preferably formed on a semiconductor integrated circuit and outputs a temperature-independent reference voltage.

[0002]

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10 ~~BACKGROUND OF THE INVENTION~~

          Up to now, there have been a large number of publications regarding a reference voltage circuit which demonstrates a temperature independent characteristic by canceling a temperature dependent characteristic and which outputs a reference voltage of the order of 1.2  
15   V.

[0003]

          First, the operation of a conventional reference voltage circuit is explained.

[0004]

20           Fig.10 shows an example of a conventional CMOS (Complementary MOS) reference voltage circuit. The reference voltage is obtained by inserting a resistor in a current loop of a reference current circuit which is termed a PTAT (Proportional to Absolute Temperature) current source circuit since in general the reference  
25   current circuit outputs the current proportional to temperature.

[0005]

In Fig.10, it is assumed that a transistor Q1 is a unit transistor and that the emitter area of a transistor Q2 is K1 times that of the unit transistor( $K1 > 1$ ).

5 [0006]

If the base width modulation is neglected, the relationship between the collector current IC to the base-to-emitter voltage VBE of a transistor is given by:

[0007]

10 
$$I_C = K \cdot I_S \exp(V_{BE}/V_T) \quad (1)$$

where IS is the saturation current of a unit transistor and VT is the thermal voltage, which is given by:

$$V_T = kT/q$$

where q is the magnitude of the unit electron charge,

15 k is Boltzmann's constant,

T is absolute temperature in kelvins, and

K is the emitter area ratio referenced to the unit transistor.

[0008]

Assuming that the DC current amplification factor of a transistor  
20 is sufficiently close to 1, and the base current is neglected, we shall find the following relationships:

$$V_{BE1} = V_T \ln\{I_{C1}/I_S\} \quad (2)$$

$$V_{BE2} = V_T \ln(I_{C2}/(K1 \cdot I_S)) \quad (3)$$

$$V_{BE1} = V_{BE2} + R1 \cdot I_{C2} \quad (4)$$

25 where  $\ln\{\}$  is a logarithmic function.

[0009]

By solving the equation (2) to (4), we obtain

$$VT \ln \{K1 \cdot IC1/IC2\} = R1 \cdot IC2 \quad (5)$$

It is noted that, since transistors Q1 and Q2 controls the common gate voltage of transistors M3 and M4 through an operational amplifier so that the equation (4) will be held valid, the transistors Q1 and Q2 are self-biased, and hence the drain currents ID3 and ID4 of the transistors M3 and M4 are equal to each other and

$$ID3 = ID4 = IC1 = IC2 \quad (6)$$

From the equation (5), we shall therefore have the following equation:

$$ID3 = ID4 = IC1 = IC2 = VT \ln(K1)/R1 \quad (7)$$

[0010]

The drain current ID3 of the transistor M3 is converted by the resistor R2 to a voltage and becomes the reference voltage VREF. That is, the reference voltage VREF is expressed as follow.

$$\begin{aligned} VREF &= VBE1 + R2 \cdot ID3 \\ &= VBE1 + R2 \cdot VT \ln(K1)/R1 \end{aligned} \quad (8)$$

[0011]

In the equation (8), the base-to-emitter voltage VBE1 of the transistor Q1, which is driven by the PTAT reference current, has a negative temperature characteristic on the order of approximately -1.9mV/°C, which is slightly less than -2mV, while the thermal voltage VT has a positive temperature characteristic of 0.0853 mV/°C.

[0012]

Accordingly, in order that the output reference voltage  $V_{REF}$  will not exhibit a temperature dependent characteristic, the cancellation of temperature dependency of the output reference voltage  $V_{REF}$  may be performed by a combination of a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic.

[0013]

That is, the value of  $(R2/R1)\ln(K1)$  is 22.3, while that of  $(R2/R1)VT\ln(K1)$  is 0.57V.

10 [0014]

If the base-to-emitter voltage  $V_{BE1}$  is 0.7 V,

$$\{V_{BE1} + (R2/R1)VT\ln(K1)\}$$

$$= 1.27V$$

[0015]

## 15 SUMMARY OF THE DISCLOSURE

Up to now, in the reference voltage circuit for outputting the reference voltage not exhibiting this sort of temperature characteristic, an operational amplifier is used in a feedback circuit, and a resistor is introduced in a current loop of the PTAT current source circuit, so that a desired resistance ratio is required. The voltage drop across a resistor on the order of approximately 0.6 V is needed for one resistor. Thus, if it is desired to diminish the driving current of a transistor connected in a diode configuration, a large resistance value is required, thus increasing a chip size.

25 [0016]

A reference voltage circuit, exemplified first and foremost by a bias voltage of the circuitry, arranged in a large number of LSIs, including digital LSIs, such as memory devices, to say nothing of an analog LSI, is routinely used. In particular, a reference voltage circuit, which outputs a voltage not exhibiting a temperature dependent characteristic, is generally termed "a band gap reference voltage circuit".

[0017]

The output voltage of a band gap reference voltage circuit is close to 1.205 V, which is the band gap voltage of Silicon at 0 ° K.

[0018]

Since the CMOS process nowadays is predominantly used, realization of a circuit with part elements that can be readily manufactured by the CMOS process has been desired. In particular, it is more desirable that a standard digital CMOS process can smoothly realize a circuit. In such case, however, a high precision resistance ratio or a high resistance leads to an increase of a chip size.

[0019]

Accordingly, it is an object of the present invention to provide a reference voltage circuit for outputting a reference voltage not exhibiting a temperature characteristic, which can be implemented only using transistors without adopting a high precision resistance ratio or a high resistance to simplify the circuit structure.

[0020]

A CMOS reference voltage circuit in accordance with one aspect

of the present invention comprises first and second diode-connected transistors(or diodes), which are grounded, and are driven respectively by two constant currents, bearing a constant current ratio, and means for amplifying a differential voltage between output voltages of the first and second diode-connected transistors(or diodes) by a predetermined constant factor and summing the resulting amplified voltage to an output voltage of the first or second diode-connected transistor(or diode), in which said means for amplification and summation includes first and second operational transconductance amplifiers (OTAs) and a current mirror circuit, in which the first OTA is fed with the differential voltage, the second OTA has a first input terminal(-) fed with an output voltage from the first or second diode-connected transistor(or diode) and a second input terminal(+) connected to an output terminal and driven with a current proportional to the output current of said first OTA, an output terminal voltage of the second OTA being an output reference voltage.

[0021]

In accordance with the present invention, the transconductance  $gm1$  of the first OTA  $gm1$  is equal to the transconductance  $gm2$  of the second OTA ( $gm1 = gm2$ ), and the current ratio of the input current to the output current in the current mirror circuit is set to  $1:K2$ , where  $K2 > 1$ , to produce a desired amplification factor.

[0022]

In accordance with the present invention, the current ratio of the input current to the output current in the current mirror circuit is equal (1:1) and the transconductance  $gm1$  of the first OTA1 and that  $gm2$  of the

second OTA 2 are set so that  $gm_1 = K_2 \times gm_2$ , where  $K_2 > 1$ , to obtain a desired amplification factor.

[0023]

In accordance with the present invention, the current ratio of the input current to the output current in the current mirror circuit is set to 1: $K_2$ , where  $K_2 > 1$ , and the transconductance  $gm_1$  of the first OTA1 and that  $gm_2$  of the second OTA 2 are set so that  $gm_1 = K_3 \times gm_2$ , where  $K_3 > 1$ , to obtain a desired amplification factor.

[0024]

A CMOS reference voltage circuit in accordance with another aspect of the present invention, comprises first and second diode-connected transistors(or diodes), which are grounded, and are driven respectively by two constant currents, bearing a constant current ratio, and means for amplifying a differential voltage between output voltages of the first and second diode-connected transistors(or diodes) by a predetermined constant factor and summing the resulting amplified voltage to an output voltage of the first or second diode-connected transistor(or diode), in which the means for amplification and summation includes  $(K_2+1)$  differential pairs,  $K_2$  being an integer not less than 1, the first differential pair being fed with the differential voltage, one of differential pair transistors of the second differential pair being fed with an output voltage of the first or second diode-connected transistor, the other of the differential pair transistors being diode-connected and being driven with the current proportional to the output current of one of the transistors of the first differential pair, output

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voltage of diode-connected transistors of the second to number K2  
differential pairs are fed to one of the differential pair transistors of the  
third to the number (K2+1) differential pairs, respectively, the other  
transistors of the differential pair transistors being diode-connected and  
5 driven by currents proportional to the output current of the one  
transistor of the first differential pair, with the first to number (K2+1)  
differential pairs being driven with the (K2+1) constant currents bearing  
a certain constant current ratio relative to one another, with the  
differential input voltages of the second to number (K2+1) differential  
10 pairs being summed together to produce a desired amplification factor.  
[0025]

A CMOS reference voltage circuit in accordance with another  
aspect of the present invention comprises first and second diode-  
connected transistors(or diodes), which are respectively grounded, and  
15 are driven by two constant currents, bearing a constant current ratio, and  
means for amplifying a differential voltage of output currents of the first  
and second diode-connected transistors(or diodes) by a preset factor and  
summing the resulting amplified voltage to an output voltage of the first  
or second diode-connected transistor(or diode), in which said means for  
20 amplification and summation includes (K2+1) differential pairs, with the  
first differential pair being fed with the differential voltage, one of  
differential pair transistors of the second differential pair being fed with  
an output voltage of the first or second diode-connected transistor, the  
other of the differential transistors being diode-connected, the  
25 differential transistors of the third to number K2 differential pairs being



diode-connected, the one diode-connected differential transistor of a given stage being driven by the constant current with the other diode-connected differential transistor of a preceding stage, the other diode-connected transistor being driven with the current proportional to the output current of the first differential pair, the first to number (K2+1) differential pairs being driven with (K2+1) constant currents bearing a certain constant current ratio to one another, with the differential input voltages of the second to number (K2+1) differential pairs being summed together to produce a desired amplification factor.

10 [0026]

A CMOS reference voltage circuit in accordance with another aspect of the present invention comprises first and second diode-connected transistors(or diodes), which are grounded, and respectively driven with two constant currents, bearing a constant current ratio, and means for amplifying a differential voltage of output voltages of the first and second diode-connected transistors(or diodes) by a preset factor and summing the resulting amplified voltage to an output voltage of the first or second diode-connected transistor(or diode), in which said means for amplification and summation is made up of two differential pairs, one of the differential transistors of a second one of the differential pairs being fed with an output voltage of the first or second diode-connected transistors(or diodes), the other differential transistor being diode-connected and being driven with the current proportional to the output current of one of the transistors of the first differential pair, the first differential pair and the second differential pair being driven with two

constant currents having a constant current ratio to each other, an operating input voltage range of the second differential pair being a constant number tuple of the operating input voltage range of the first differential pair to produce a desired amplification factor.

5 [0027]

In accordance with the present invention, the first diode-connected transistor (or diode) is equal to the second diode-connected transistor (or diode), with the ratio of respective driving currents not being equal to 1.

10 [0028]

In accordance with the present invention, the size of the first diode-connected transistor (or diode) is  $K_1$  times the size of the second diode-connected transistor (or diode), with the driving current ratio not being equal to 1.

15 [0029]

In accordance with the present invention, the size of the first diode-connected transistor (or diode) differs from the size of the second diode-connected transistor (or diode), with the driving current ratio being equal to 1.

20 [0030]

In accordance with the present invention, the gate  $W/L$  ratio of each transistor of the first differential pair is  $K_2$  times the gate  $W/L$  ratio of each transistor of the second differential pair,  $W$  and  $L$  being the gate width and the gate length of the transistor, respectively, the driving  
25 current of the second differential pair being  $K_3$  times the driving current

of the third differential pair, the output current of the first differential pair being multiplied by  $K_3$  to drive the diode-connected transistor of the second differential pair to produce the desired amplification factor.

[0031]

5           The present invention is constituted by a grounded diode-connected transistor, (or diode) driven at a constant current, and an operational amplifier having a voltage follower type offset, for receiving an output voltage of the diode-connected transistor (or diode).

[0032]

10           In accordance with the present invention, the operational  
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 amplifier is driven with the constant current, each of two transistors making up an input differential pair has a gate W/L ratio of  $1:K_2$ , and the gate W/L ratio of the two transistors forming an active load operating as a load to the two transistors is  $K_3:1$ , with offset values being summed  
 15 together.

[0033]

          In accordance with the present invention, each of two transistors making up an input differential pair has a gate W/L ratio of  $K_2:1$ , and the gate W/L ratio of the two transistors forming an active load operating as  
 20 a load to the two transistors is  $1:K_3$ , with offset values being subtracted.

          Still other objects and advantages of the present invention will become readily apparent to those skilled in this art from the following detailed description in conjunction with the accompanying drawings wherein only the preferred embodiments of the invention are shown and  
 25 described, simply by way of illustration of the best mode contemplated

of carrying out this invention. As will be realized, the invention is capable of other and different embodiments, and its several details are capable of modifications in various obvious respects, all without departing from the invention. Accordingly, the drawing and description  
5 are to be regarded as illustrative in nature, and not as restrictive.

#### BRIEF DESCRIPTION OF THE DRAWINGS

Fig.1 shows a circuit structure of an embodiment of the present invention;

Fig.2 illustrates the operation of multiplication of a reference  
10 voltage circuit embodying the present invention;

Fig.3 illustrates the operation of multiplication of the reference voltage circuit embodying the present invention;

Fig.4 illustrates the operation of multiplication of the reference voltage circuit embodying the present invention;

15 Fig.5 shows a circuit structure of a second embodiment of the present invention;

Fig.6 shows a circuit structure of a third embodiment of the present invention;

20 Fig.7 shows a circuit structure of a fourth embodiment of the present invention;

Fig.8 shows a circuit structure of a fifth embodiment of the present invention;

Fig.9 shows a modification of the fifth embodiment of the present invention; and

25 Fig.10 shows the structure of a reference voltage circuit

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employing a conventional operational amplifier.

[0034]

#### PREFERRED EMBODIMENTS OF THE INVENTION

The preferred embodiments of the present invention will now be  
5 described. If two transistors, both of which have emitters grounded and  
are connected in diode configuration, are driven by a current mirror  
circuit, and the current densities of the two transistors are made different  
from each other to produce different base-to-emitter voltages  $V_{BE}$ s, a  
differential voltage  $\Delta V_{BE}$  of the base-to-emitter voltages of the two  
10 transistors is proportional to absolute temperature, and hence a voltage  
proportional to the thermal voltage  $V_T$  is obtained.

[0035]

On the other hand, the base-to-emitter voltage  $V_{BE}$  of a transistor  
has a negative temperature characteristic on the order of approximately  
15  $-2 \text{ mV}/^\circ\text{C}$  to  $-1.9 \text{ mV}/^\circ\text{C}$ .

[0036]

In general, in a conventional reference voltage circuit, a  
reference voltage circuit which outputs a constant voltage not exhibiting  
a temperature dependent characteristic, is realized by weight-summing  
20 the voltage  $V_{TPAT}$  proportional to absolute temperature and the voltage  
 $V_{IPTAT}$  inversely proportional to absolute temperature.

[0037]

This constant voltage is of a voltage value such that  
 $V_{PTAT} + V_{IPTAT} \cong 1.2\text{V}$  or thereabout.

25 [0038]

In the conventional reference voltage circuit, this weight summation of the voltages VPTAT and VIPTAT is realized by a resistor inserted on a PTAT current path of VIPTAT, and is termed " $\Delta V$  multiplier".

5 [0039]

In accordance with the present invention, the weight summation is not realized by a resistor, but by a differential pair.

[0040]

10 SUB  
A3 7 An embodiment of the present invention is shown in Fig.1, showing two OTAs (operational transconductance amplifiers), in which a differential input voltage is proportional to an output current and in which a transconductance exhibits a linear characteristic. Across these OTAs, the current ( $K2 \times g_{m1} \Delta V_{BE}$ ) having a predetermined constant ratio K2 to an output current ( $g_{m1} \Delta V_{BE}$ ) of the first OTA 11, which is

15 proportional to a differential voltage  $V_{BE}$  ( $=V_{BE2} - V_{BE1}$ ) of the base-to-emitter voltage  $V_{BE}$  of two bipolar transistors Q1 and Q2, is caused to flow into the second OTA 12 to produce a voltage corresponding to the differential voltage  $\Delta V_{BE}$  multiplied by a constant value, that is VPTAT ( $= K2 \times g_{m1} \Delta V_{BE}/g_{m2}$ ). In the

20 second OTA 12, the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2 is summed to VPTAT and the resulting voltage is output to produce a desired constant voltage  $V_{REF}$  not exhibiting a temperature dependent characteristic.

[0041]

25 In a modification of the present invention, as shown in Figs.5 and

6, plural differential pairs are connected in a cascaded configuration so that differential voltages applied to the differential input terminals of the respective differential pairs will be made equal to each other and equal to the differential voltage  $\Delta V$ . From the differential pair of the last stage, a voltage equal to an integer number multiple of the differential voltage  $\Delta V$  is obtained as a voltage proportional to absolute temperature.

[0042]

Alternatively, a transfer curve (transfer characteristic) of the differential pairs may be normalized by a square root of a ratio of the driving current  $I_0$  to a transconductance parameter  $\beta$  of the differential transistor, or  $\sqrt{I_0/\beta}$ , and thus may be constant, as shown in Fig.7

[0043]

That is, if a normalized current equal to the normalized current flowing through one transistor of the first differential pairs M1 and M2 by the voltage applied to the first differential pairs M1 and M2 is caused to flow through one transistor of the second differential pair M3 and M4, the voltage across input terminals of the second differential pair is multiplied with the ratio of the standardized voltages of the two differential pairs, or divided by the ratio of the standardized voltages of the two differential pairs if the ratio is less than 1.

[0044]

Therefore, the summation may be made as the voltage applied to the other input terminal of the second differential pair is multiplied with the voltage applied across the input terminals of the first differential

pair.

[0045]

Alternatively, an offset voltage VOS, generated in a voltage follower circuit, made up of an unbalanced differential pair, including unbalanced input differential pairs M1 and M2, active load devices M3 and M4, an output stage M5 and phase compensation circuits RC and CC, is obtained as a voltage VPTAT which is proportional to absolute temperature. In the operational amplifier which is driven with a constant current, the gate W/L ratio (gate width/gate length ratio) of the two transistors M1 and M2 constituting the input differential pair is 1:K2, and the gate W/L ratio of two transistors M3 and M4 (of a current mirror circuit configuration) forming an active load operating respectively as a load for two transistors M1 and M2, is K3:1, and the offset is added to an input voltage to the voltage follower circuit to produce an output reference voltage VREF. Alternatively, the gate W/L ratio of the two transistors forming the input differential pair is K2:1, while the gate W/L ratio of the two transistors forming an active load operating as a load for the two transistors is 1:K3, and an offset is subtracted from an input voltage to produce an output reference voltage VREF.

20 [0046]

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AL 7  
~~Such a configuration may also be used which includes a source-grounded MOS transistor M10, having its drain and gate connected to one end and the other end of the resistor R1, respectively, a source-grounded MOS transistor M11 having its gate connected to the drain of the MOS transistor M11 and a current mirror circuit, having its input end~~



connected to a drain of the MOS transistor M11, and adapted for supplying the constant current to the MOS transistor M10, a common source of the first and second MOS transistors M1 and M2 of the differential pair, a MOS transistor M5 of a source follower configuration and to the collector of the bipolar transistor Q1.

[0047]

Referring to the drawings, certain preferred embodiments of the present invention are explained in detail. Fig.1 shows a circuit configuration of an embodiment of the present invention, as applied to a CMOS reference voltage circuit. As shown in Fig.1, this circuit includes first and second emitter-grounded transistors Q1 and Q2, each of which has a base connected to a collector and is provided with a constant current at the collector, first and second operational transconductance amplifiers (abbreviate to OTAs) 11 and 12, each of which outputs current corresponding to the voltage difference between the voltage at a positive phase(non-inverting) input terminal (+) and that at a reverse phase(inverting) input terminal (-), and a current mirror circuit 13 which has a ratio of the current input to the input end to the current output from the output end equal to a predetermined value K2. The reverse phase input terminal (-) and the positive phase input terminal (+) of the first OTA 11 are connected to the collectors(more precisely the connection nodes of the collectors and the bases) of the first and second transistors Q and Q2, respectively. The first OTA 11 has its output terminal connected to an input end of the current mirror circuit 13. An output end of the current mirror circuit 13 and the

SUB A5 7 collector of the second transistor Q2 are connected to the positive phase input terminal (+) and the reverse phase input terminal (-) of the second OTA 12, respectively, while the output terminal of the second OTA 12 is connected to the positive phase input terminal (+) of the second OTA 12.

5 The reference voltage VREF is output at an output terminal of the second OTA 12.

[0048]

SUB A6 7 It is assumed that, in the embodiment shown in Fig.1, the emitter area of the transistor Q1 is K1 times the emitter area of the transistor Q2.

10 The collectors of the transistors Q1 and Q2 are connected to drains of the P-channel MOS transistors M2 and M3, that is output terminals of a current mirror circuit(made up of P-channel MOS transistors M1, M2 and M3) which receives a constant current I0 from a constant current source 14 at its input terminal, and the current I0 flows through collectors of

15 the transistors Q1 and Q2.

[0049]

SUB A7 7 If the DC current amplification factor of the transistors is sufficiently close to unity and the base current is neglected, from the above equation (1), the base-to-emitter voltages VBE and VBE2 of the

20 transistors Q1 and Q2 are expressed as follows:

$$\begin{aligned} V_{BE1} &= V_T \ln \{ I_{C1} / (K1 \cdot I_S) \} \\ &= V_T \ln \{ I_0 / (K1 \cdot I_S) \} \end{aligned} \quad (9)$$

$$\begin{aligned} V_{BE2} &= V_T \ln (I_{C2} / I_S) \\ &= V_T \ln (I_0 / I_S) \end{aligned} \quad (10)$$

25 [0050]

The differential voltage  $\Delta V_{BE}$  between the base-to-emitter voltages  $V_{BE1}$  and  $V_{BE2}$  is given by:

$$\begin{aligned}\Delta V_{BE} &= V_{BE2} - V_{BE1} \\ &= V_T \ln(K_1)\end{aligned}\quad (11)$$

5 [0051]

That is, if the emitter grounded two transistors Q1 and Q2, both connected in diode configuration, are driven by the current mirror circuit, the densities of currents which flow respectively through the two transistors are rendered different as are the base-to-emitter voltages, and  
10 the differential voltage  $\Delta V_{BE}$  between the base-to-emitter voltages of the two transistors Q1 and Q2 is taken, the differential voltage  $\Delta V_{BE}$  is proportional to absolute temperature, thus producing a voltage proportional to the thermal voltage  $V_T$ .

[0052]

15 Also, as may be seen from the equation (12), in order to set current densities of two transistors different to produce a voltage differential between the base-to-emitter voltages of the two transistors Q1 and Q2, any of the following methods may be effectively used:

the emitter areas of the two transistors Q1 and Q2 are rendered  
20 different, as the driving currents supplied to the collectors of the two transistors Q1 and Q2 are kept equal;

the emitter areas of the two transistors Q1 and Q2 are rendered equal to each other and the driving currents supplied to the collectors of the two transistors Q1 and Q2 are rendered different; or

25 both the driving currents and emitter areas of the two transistors

Q1 and Q2 are rendered different.

[0053]

Next, a multiplication-summation circuit, employing two OTASs, is described.

5 [0054]

The first OTA 11 has a transconductance  $gm1$  and receives the differential voltage  $V_{BE}$  to draw the current  $gm1 \Delta V_{BE}$ . The second OTA 2 has a transconductance  $gm2$  and has a reverse phase input terminal (-) for receiving the base-to-emitter voltage  $V_{BE2}$  of one of the transistors, while having a positive phase input terminal (+) connected in common with its output terminal and driven with a current  $K2 \times gm1 \Delta V_{BE}$  through the current mirror circuit 13.

[0055]

In order for the two OTAs to have the function of the voltage multiplication circuit, the OTAs need to be equal in transconductance ( $gm1 = gm2$ ), as shown in Fig.2. If the current ratio (that is, input current value: mirror current value) of the current mirror circuit 13 is set to 1:K2 ( $K2 > 1$ ), the voltage gain is K2 and, since the output current of the second OTA 12 is

$$20 \quad K2 \times gm1 \Delta V_{BE} \quad (12),$$

the input differential voltage of the second OTA 12 is the output current divided by transconductance  $gm2$ , so that we have

$$\begin{aligned} \Delta V &= K2 \cdot gm1 \Delta V_{BE} / gm2 \\ &= K2 \Delta V_{BE} \end{aligned} \quad (13)$$

25 [0056]

Since in the second OTA, the output terminal for outputting the reference voltage  $V_{REF}$ , is connected to the positive phase input terminal (+), the voltage of the reverse phase input terminal (-) is  $V_{BE2}$  and  $\Delta V = (V_{REF} - V_{BE2})$ , the reference voltage  $V_{REF}$  is given by:

$$\begin{aligned} V_{REF} &= V_{BE2} + K2 \Delta V_{BE} \\ &= V_{BE2} + K2 \cdot V_T \ln(K1) \end{aligned} \quad (14)$$

[0057]

In the equation (14), the base-to-emitter voltage  $V_{BE2}$  of the transistor  $Q2$ , driven with the constant current  $I_0$ , has a negative temperature characteristic on the order of approximately  $-2 \text{ mV}/^\circ\text{C}$ , while the thermal voltage  $V_T$  has a positive temperature characteristic on the order of approximately  $0.0853 \text{ mV}/^\circ\text{C}$ .

[0058]

Thus, in order for the output reference voltage  $V_{REF}$  not to exhibit a temperature dependent characteristic, the temperature dependent characteristic might be cancelled with a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic.

[0059]

That is, the value of  $K2 \ln(K1)$  is 23.45, with the value of  $K2 \cdot V_T \ln(K1)$  being 0.61V. If  $V_{BE2}$  is 0.7V,

$$\{V_{BE2} + K2 \cdot V_T \ln(K1)\} = 1.31 \text{ V.}$$

[0060]

Alternatively, in order for these two OTA to have the function of a voltage multiplication circuit, as shown in Fig.3, it is also sufficient if

the two conductance values differ from each other such that

$$g_{m1} = K_2 g_{m2} (K_2 > 1)$$

and, when the current ratio of the current mirror circuit is set to 1:1, the voltage gain is  $K_2$ , such that a differential voltage  $K_2 \Delta V$  is obtained as an output voltage:

$$\begin{aligned} K_2 \Delta V &= g_{m1} \Delta V_{BE} / g_{m2} \\ &= K_2 \Delta V_{BE} \end{aligned} \quad (15)$$

Thus we have

$$\begin{aligned} V_{REF} &= V_{BE2} + K_2 \Delta V_{BE} \\ &= V_{BE2} + K_2 \cdot V_T \ln(K_1) \end{aligned} \quad (16)$$

[0061]

In the equation (16), the base-to-emitter voltage  $V_{BE2}$  of the transistor  $Q_2$ , driven with the constant current  $I_0$ , has a negative temperature characteristic on the order of approximately  $-2\text{mV}/^\circ\text{C}$ , while the thermal voltage  $V_T$  has a positive temperature characteristic on the order of approximately  $0.0853\text{ mV}/^\circ\text{C}$ . Thus, in order for the output reference voltage  $V_{REF}$  not to exhibit a temperature dependent characteristic, the temperature dependent characteristic might be cancelled with a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic.

[0062]

That is, the value of  $K_2 \ln(K_1)$  is 23.45, with the value of  $K_2 \cdot V_T \ln(K_1)$  being 0.61V. If  $V_{BE2}$  is 0.7V, we have

$$\{V_{BE2} + K_2 \cdot V_T \ln(K_1)\} = 1.31\text{V}.$$

[0063]

Alternatively, in order for the two OTA to have the function of a voltage multiplication circuit, as shown in Fig.4, it is also sufficient if the two conductance values differ from each other, as shown in Fig.4, such that

$$5 \quad g_{m1} = K_3 g_{m2} (K_3 > 1)$$

and, when the current ratio of the current mirror circuit is set to 1:K<sub>2</sub>, the voltage gain is K<sub>4</sub>, such that a differential voltage  $K_4 \Delta V$

$$\begin{aligned} K_4 \Delta V &= K_2 g_{m1} \Delta V_{BE} / g_{m2} \\ &= K_2 \cdot K_3 \Delta V_{BE} \end{aligned} \quad (17)$$

10 is obtained as the output voltage.

[0064]

Thus, we have

$$\begin{aligned} V_{REF} &= V_{BE2} + K_4 \Delta V_{BE} \\ &= V_{BE2} + K_2 \cdot K_3 \cdot V_T \ln(K_1) \end{aligned} \quad (18).$$

15 [0065]

In the equation (18), the base-to-emitter voltage  $V_{BE2}$  of the transistor Q<sub>2</sub>, driven with the constant current  $I_0$ , has a negative temperature characteristic on the order of approximately  $-2\text{mV}/^\circ\text{C}$ , while the thermal voltage  $V_T$  has a positive temperature characteristic on the order of approximately  $0.0853\text{ mV}/^\circ\text{C}$ . Thus, in order for the output reference voltage  $V_{REF}$  not to exhibit a temperature dependent characteristic, the temperature dependent characteristic might be canceled by a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic.

25 [0066]

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That is, the value of  $K2 \cdot K3 \ln(K1)$  is 23.45, with the value of  $K2 \cdot K3 \cdot VT \ln(K1)$  being 0.61V. If  $V_{BE2}$  is 0.7V, we have

$$\{V_{BE2} + K2 \cdot K3 \cdot VT \ln(K1)\} = 1.31V.$$

[0067]

5           The second embodiment of the present invention will now be described. Fig.5 shows a circuit configuration of a second embodiment of the CMOS reference voltage circuit according to the present invention. Referring to Fig.5, this embodiment includes first and second diode-connected common emitter transistors Q1 and Q2, driven by two constant  
10           currents having a constant current ratio. The present embodiment also includes  $(K2+1)$  differential pairs as means for amplifying the differential voltage of the output voltages of the transistors Q1 and Q2 by a preset factor and summing the so amplified differential voltage to the output voltage (collector voltage) of the transistors Q1 and Q2.

15           [0068]

          The first differential pair M1 and M2 receives as a differential input a differential voltage of output voltages of the transistors Q1 and Q2.

[0069]

20           An output voltage (collector voltage) of the transistor Q2 is applied to the gate of one transistors M3 of the second differential pair M3 and M4, with the other transistor M4 of the second differential pair being connected in a diode configuration. The second differential pair is driven with the current proportional to the output current of the one  
25           transistor M2 of the first differential pair.



[0070]

In the third to number  $(K2+1)$  differential pairs, output voltages of the other diode-connected transistors  $M4$  to  $M(2K2)$  of the preceding second to number  $K2$  differential pairs are fed to the gates of the one transistors of the third to number  $(K2+1)$  differential pairs. These other transistors of the differential pairs are connected in diode configuration and are driven with the current proportional to the output current of the one transistor  $M2$  of the first differential pair.

[0071]

The first to number  $(K2+1)$  differential pairs are driven with constant  $(K2+1)$  currents respectively.

[0072]

The output voltage of the diode-connected transistor  $M(2K2+2)$ , of the transistors of the number  $(K2+1)$  differential pair, is output as the reference voltage  $VREF$ . The differential input voltages of the second to number  $(K2+1)$  differential pairs are summed together to attain a desired amplification factor.

[0073]

Referring to Fig.5,  $(K2+4)$  pieces of P-channel MOS transistors  $MP1$  and  $MP2$  to  $MP(K2+4)$ , having sources connected in common to a power supply  $VDD$  and having gates connected in common, form a first mirror circuit having  $(K2+3)$  outputs. The P-channel MOS transistor  $MP1$ , the drain of which is connected to its gate has its drain connected to a constant current source 15. A constant current  $I0$  is the input current to the first current mirror circuit. From the drains of the P-channel

MOS transistors MP2 and MP3, the constant current is fed to the collectors of the first and second transistors Q1 and Q2, whereas, from the drains of the P-channel MOS transistors MP4 to MP(K2+4), the constant current is supplied to the sources of the first to number (K2+1) differential pairs. An N-channel MOS transistor MN01, having a source grounded, a drain connected to its gate and having the drain connected to the transistor M2, and N-channel MOS transistors MN02, MN03, ... and MN0(K2+1) having sources grounded and having gates connected in common to the gate of the transistor MN01, make up a second current mirror circuit.

[0074]

The gates of the transistors M1 and M2, forming a first differential pair, are supplied with the differential voltage  $\Delta V_{BE}$  between the base-to-emitter voltages  $V_{BE1}$  and  $V_{BE2}$  of the first and second diode-connected common emitter transistors Q1 and Q2. The drain of the transistor M1 is grounded, while the drain of the transistor M2 is connected to the drain of the N-channel MOS transistor MN01 which forms an input end of the second current mirror circuit.

[0075]

Regarding transistors M3 and M4, which forms the second differential pair, the gate of the transistor M3 is connected to the collector of the transistor Q2, and hence fed with the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2. The gate and the drain of the other transistor M4 are connected in common (in a diode configuration) and connected to the drain of the N-channel MOS transistor MN02, while the

transistor M4 is driven with a current proportional to the current flowing through the other transistor M2 forming the first differential pair. The input differential voltage, applied to the gates of the transistors M3 and M4 of the second differential pair, is equal to the input differential voltage applied to the gates of the transistors M1 and M2 of the first differential pair, and is equal to  $\Delta V_{BE}$ . The gate voltage of the MOS transistor M4 is the gate voltage of the MOS transistor M3 (the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2) summed with  $\Delta V_{BE}$ , this voltage ( $V_{BE2} + \Delta V_{BE}$ ) being also fed to the gate of the one transistor M5 of the third differential pair. The other transistor M6 of the third differential pair is connected in a diode configuration and is connected to the drain of the output transistor ~~M4~~<sup>MN03</sup> of the second current mirror circuit. The differential voltage, fed to the gates of the transistors M5 and M6, is equal to the input differential voltage, applied to the gates of the transistors M1 and M2 of the first differential pair, and is equal to  $\Delta V_{BE}$ . The gate voltage of the transistor M6 is the gate voltage of the transistor M5 ( $V_{BE2} + \Delta V_{BE}$ ) summed with  $\Delta V_{BE}$  ( $V_{BE2} + 2\Delta V_{BE}$ ), this voltage being input to the gate of one transistor of a fourth differential pair, not shown. The same holds for the third to number (K2+1) differential pairs. In each of the third to number (K2+1) differential pair, the differential input voltage is equal and the output voltage ~~of~~ is higher by  $\Delta V_{BE}$  than the output voltage of the differential pair of the previous stage. The output voltage of the diode-connected other transistor of the number n stage differential pair is

25 is

$$V_{BE2} + (n-1) \times \Delta V_{BE}.$$

The reference voltage  $V_{REF}$ , which is an output voltage of the other diode-connected transistor  $M$  ( $2K_2+2$ ) of the number ( $K_2+1$ ) differential pair, is given by:

$$V_{BE2} + K_2 \times \Delta V_{BE}.$$

[0076]

Thus,

$$V_{REF} = V_{BE2} + K_2 \Delta V_{BE}$$

$$= V_{BE2} + K_2 \cdot V_T \ln(K_1) \quad (19).$$

10 [0077]

In the equation (19), the base-to-emitter voltage  $V_{BE2}$  of the transistor  $Q_2$ , driven with the constant current  $I_0$ , has a negative temperature characteristic on the order of approximately  $-2\text{mV}/^\circ\text{C}$ , while the thermal voltage  $V_T$  has a positive temperature characteristic on the order of approximately  $0.0853\text{ mV}/^\circ\text{C}$ .

[0078]

Thus, in order for the output reference voltage  $V_{REF}$  not to exhibit a temperature characteristic, it is sufficient if a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic cancels the temperature characteristic.

20 [0079]

That is, the value of  $K_2 \ln(K_1)$  is 23.45, with the value of  $K_2 \cdot V_T \ln(K_1)$  being 0.61V.

25 If  $V_{BE2}$  is 0.7V,

$$\{V_{BE2} + K2 \cdot V_T \ln(K1)\} = 1.31V.$$

[0080]

A third embodiment of the present invention is explained. Fig.6 is a circuit diagram showing a third embodiment of the CMOS reference voltage circuit according to the present invention. Referring to Fig.6, the present embodiment includes first and second diode-connected transistors Q1 and Q2, respectively grounded and driven by two constant currents having a constant current ratio to each other. As means for amplifying the differential voltage of the transistors Q1 and Q2 by a certain factor and summing the resulting amplified differential voltage to an output voltage of the first or second diode-connected transistor Q1 or Q2, (K2+1) differential pairs are arranged.

[0081]

The first differential pair M1 and M2 is fed with a differential voltage of the output voltages of the transistors Q1 and Q2, while the second differential pair M3 and M4 is configured so that an output voltage of the transistor Q2 is applied to the one transistor M3 of the differential pair, with the other transistor M4 thereof being connected in a diode configuration.

[0082]

The differential transistors M5 and M6 to M(2K2-1) and M(2K2) of the third to number K2 differential pairs are all connected in a diode configuration, in such a manner that a diode-connected transistor of the differential pair of the front stage and the diode-connected transistor of the differential pair of the back side stage are driven at a constant

current with a constant current ratio  $K_2$ , while the number  $M(2K+1)$  and  $M(2K+2)$  transistors of the number  $(K_2+1)$  differential pair are all connected in a diode configuration. The diode connected number  $M(2K_2 + 1)$  transistor is driven at a constant current along with the diode-connected transistor  $M(2K_2)$  of the front stage, while the diode-connected other transistor  $M(2K_2+2)$  is driven with the current proportional to one output current of the first differential pair

[0083]

The number 1 to number  $(K_2+1)$  differential pairs is driven at a constant current  $(K_2+1)$  with a constant current ratio. The desired amplification factor is achieved by summing all of the differential input voltages of the second to number  $(K_2+1)$  differential pairs.

[0084]

Referring to Fig.6,  $(K_2+4)$  common gate P-channel MOS transistors  $MP_1, MP_2$  to  $MP(K_2+4)$ , having respective sources connected to a power supply  $V_{DD}$  in common, constitute a first current mirror circuit having  $(K_2+3)$  outputs. The P-channel MOS transistor  $MP_1$ , having its drain connected to its gate has the drain connected to a constant current source 16 with a constant current  $I_0$  being an input current to the first current mirror circuit. The drains of the P-channel MOS transistors  $MP_2$  and  $MP_3$  output a constant current to the collectors of the first and second transistors  $Q_1$  and  $Q_2$ , while the drains of the P-channel MOS transistors  $MP_4$  to  $MP(K_2+4)3$  output a constant current to the common source of the number 1 to number  $K_2+1$  differential pairs.

The transistor  $MN_{01}$ , having its source grounded, having its drain

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connected to its gate, having the drain connected to the constant current source IO and fed with the sink current, and the N-channel MOS transistors MN04, MN05 and MN(K2+1), having the sources grounded and having the gates to the gate of the transistor MN01 in common, constitute a second current mirror circuit. The transistor MN02, having its source grounded, having its drain connected to its gate, and having the drain connected to the drain of the transistor MN02, and an N-channel MOS transistor MN03, having its source grounded and having its gate connected to the gate of the transistor MN02 in common, constitute a third current mirror circuit.

[0085]

In Fig.6, the first differential pair, made up of P-channel MOS transistors M1 and M2, is fed with a differential input voltage corresponding to a differential voltage  $\Delta V_{BE}$  between the base-to-emitter voltage  $V_{BE1}$  of the transistor Q1 and the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2.

[0086]

Also, in the second differential pair, made up of the transistors M3 and M4, the transistor M3 has its gate fed with the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2, while the transistor M4 has its gate connected to its drain in a diode configuration and also connected to a gate of the transistor M5 which constitutes a third differential pair and has its gate connected to its drain in a diode configuration. The transistor M4 and M5 are driven with a constant current.

[0087]

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 The third to number K2 differential pairs are configured in similar manner. The diode-connected transistor M (2K+2) of the number K2+1 last-stage differential pair has its drain connected to the drain of the output transistor MN03 of the third current mirror circuit, and is driven by the current proportional to that flowing in the transistor M2 of the first differential pair.

[0088]

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 The first differential pair (made up of transistors M1 and M2) is driven by the transistor MP4 with a current  $I_0$  proportional to the constant current  $I_0$ . If a differential voltage  $\Delta V_{BE}$  is differentially input to the first differential pair, and the drain currents flowing through the transistors M1 and M2 of the first differential pair are  $I_1$  and  $I_2$ , we have

$$I_1 + I_2 = I_0.$$

10 The common source of the transistors MP (2K+1) and MP(2K2+2) of the last-stage number (K2+1) differential pair is fed with the current  $I_0$  from the transistor MP(K2+4), the drain of the transistor M(2K2+2) is driven by the transistor MN03 with the current  $I_2$ , with the current  $I_0 - I_2 = I_1$  flowing through the drain of the transistor M(2K2+1). The differential input voltage of the number (K2+1) is  $\Delta V_{BE}$ , with the gate voltage of the transistor M(2K2+1) being lower by  $\Delta V_{BE}$  than the gate voltage of the transistor M(2K2+2).

[0089]

The drains of the transistor M(2K2+1) of the number (K2+1) differential pair and the drain of the transistor M(2K2) of the number



(K2) differential pair are connected in common to the drain of the output transistor MN0(K2+3) of the second current mirror circuit. Since these transistors are driven with the current  $I_0$  proportional to the constant current  $I_0$ , the current flowing through the drain of the transistor MP(2K2) is  $I_0 - I_1 = I_2$ , while the current flowing through the drain of the transistor MP(2K2-1) is  $I_0 - I_2 = I_1$ . As in the case of the first differential pair, the differential input voltage is  $\Delta V_{BE}$ , with the gate voltage of the transistor MP(2K2-1) being lower by  $\Delta V_{BE}$  than the gate voltage of the transistor M(2K2). In this manner, the gate voltages of the diode-connected transistors are lowered stepwise by  $\Delta V_{BE}$  up to the second differential pair M3 and M4.

[0090]

Since the voltage input to the gate of the transistor M3 of the second differential pair is the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2, the drain (gate) voltage of the transistor M4 of the second differential pair is  $V_{BE2} + \Delta V_{BE}$ , so that the output voltage  $V_{REF}$  of the transistor M(2K2+2) of the number (K2+1) stage differential pair is

$$\begin{aligned} V_{REF} &= V_{BE2} + K2 \Delta V_{BE} \\ &= V_{BE2} + K2 \cdot V_T \ln(K1) \end{aligned} \quad (20).$$

[0091]

In the equation (20), the base-to-emitter voltage  $V_{BE2}$  of the transistor Q2, driven with the constant current  $I_0$ , has a negative temperature characteristic on the order of approximately  $-2\text{mV}/^\circ\text{C}$ , while the thermal voltage  $V_T$  has a positive temperature characteristic of  $0.0853\text{ mV}/^\circ\text{C}$ .

[0092]

Therefore, in order that the output reference voltage  $V_{REF}$  will not exhibit a temperature characteristic, it is sufficient if a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic ekes out the temperature characteristic. That is, the value of  $K2 \ln(K1)$  is 23.45, while that of  $K2 \cdot V_T \ln(K1)$  is 0.57V.

If  $V_{BE2}$  is 0.7 V,

$$\{V_{BE2} + K2 \cdot V_T \ln(K1)\} = 1.31V.$$

10 [0093]

A fourth embodiment of the present invention is now explained. Fig.7 shows a circuit structure of a fourth embodiment of a CMOS reference current circuit of the present invention. In the present embodiment, a multiplication summation circuit employing two differential pairs is described.

[0094]

The embodiment shown in Fig.7 includes first and second diode-connected transistors  $Q1$  and  $Q2$ , both of which have emitters, grounded and are driven with two constant currents bearing a constant current ratio to each other. The present embodiment also includes means, made up of two differential pairs, for amplifying the differential voltage of the output voltages of the two transistors  $Q1$  and  $Q2$  by a preset constant factor and summing the result to an output voltage from the transistor  $Q2$ .

25 [0095]

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The differential pair made up of P-channel MOS transistors M1 and M2 receives a differential voltage of output voltages of the transistors Q1 and Q2. An output voltage of the transistor Q2 is applied to the gate of the P-channel MOS transistor M3 making up a second differential pair along with the diode-connected P-channel MOS transistor M4. The drain of the transistor M4 is driven with a current proportional to an output current of the first differential pair, that is the drain current of the transistor M2 (K3 tuple current). The common source of the first and second differential pairs is driven with two constant currents, having a certain current ratio to each other. A desired amplification factor is realized by setting the operating input voltage range of the second differential pair so as to be a preset constant number tuple of that of the first differential pair. In Fig.7, P-channel MOS transistors M5 to M9, which have their sources connected in common to a power supply VDD and have their gates connected in common, constitute a first current mirror circuit. The P-channel MOS transistor M9, having its drain connected to its gate has the drain connected to a constant current source 17, with the constant current I0 being an input current to the current mirror circuit. From the drains of the P-channel MOS transistors M5 and M7, constant currents are fed to the collectors of the first and second transistors Q1 and Q2, and from the drains of the P-channel MOS transistors M6 and M8, constant currents are fed to the commonly connected sources of the first and second differential pairs respectively. An N-channel MOS transistor MN10, having a source grounded, having a drain and a gate connected to each

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Sub All other and having the drain connected to the drain of the transistor M2, and an N-channel MOS transistor MN11, having a source grounded and having a gate connected to the gate of the transistor MN10, constitute a second current mirror circuit.

5 [0096]

It is assumed that a transconductance parameter  $\beta$  is equal for both the transistors M1 and M2, forming the first differential pair, and that the two transistors are driven at the constant current  $I_0$ , where the transconductance parameter  $\beta$  is given by:

10 
$$\beta = \mu (C_{ox}/2)(W/L)$$

where  $\mu$  is the effective mobility of the carrier,  $C_{ox}$  is the capacity of a gate oxide film per unit area and  $W$  and  $L$  are a gate width and a gate length, respectively. The gate  $W/L$  ratio of the transistors M1 and M2, forming the first differential pair, where  $W$  and  $L$  denote the gate width and the gate length, respectively, is set so as to be  $K_2$  times the gate  $W/L$  ratio of the transistors M3 and M4 forming the second differential pair.

[0097]

The respective drain currents  $I_{D1}$  and  $I_{D2}$  of the transistors M1 and M2 are given by:

20 
$$I_{D1} = \beta (V_{GS1} - V_{TH})^2 \quad (21)$$

$$I_{D2} = \beta (V_{GS2} - V_{TH})^2 \quad (22),$$

respectively. In the above equations,  $V_{GS1}$  and  $V_{GS2}$  denote the gate-to-source voltages of the transistors M1 and M2 respectively and  $V_{TH}$  is a threshold voltage.

25 [0098]

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The sources of the transistors M1 and M2 are connected to the drain of the P-channel MOS transistor M6, forming an output of the first current mirror circuit. From the conditions for the driving current,

$$I_{D1} + I_{D2} = I_0 \quad (23)$$

5 [0099]

By solving equation (21) to (23),  $I_{D1}$  and  $I_{D2}$  are given by the following equation (24) and (25):

[0100]

$$I_{D1} = \frac{1}{2} \{ I_0 + \beta \Delta V \sqrt{\frac{2I_0}{\beta} - (\Delta V)^2} \} \quad (24)$$

[0101]

$$I_{D2} = \frac{1}{2} \{ I_0 - \beta \Delta V \sqrt{\frac{2I_0}{\beta} - (\Delta V)^2} \} \quad (25)$$

10 In the above equations,  $\Delta V = V_{GS1} - V_{GS2}$ .

[0102]

If the equation (24) and (25) are normalized with the current  $I_0$ , the following equations (26) and (27) are obtained:

[0103]

$$f_{D1}(x) = \frac{I_{D1}}{I_0} = \frac{1}{2} \{ 1 + \sqrt{2} \sqrt{1 - \frac{x^2}{2}} \} \quad (26)$$

15

[0104]

$$f_{D2}(x) = \frac{I_{D2}}{I_0} = \frac{1}{2} \{ 1 - \sqrt{2} \sqrt{1 - \frac{x^2}{2}} \} \quad (27)$$

[0105]

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where

[0106]

$$x = \Delta V / \sqrt{I_0 / (K_2 \beta)}$$

[0107]

In the second differential pair, made up by the transistors M3 and

5 M4, the above equations are given by:

[0108]

$$f_{D3}(x) = \frac{I_{D3}}{K_3 I_0} = \frac{1}{2} \left\{ 1 + \sqrt{2} \sqrt{1 - \frac{x^2}{2}} \right\} \quad (28)$$

and

[0109]

$$f_{D4}(x) = \frac{I_{D4}}{K_3 I_0} = \frac{1}{2} \left\{ 1 - \sqrt{2} \sqrt{1 - \frac{x^2}{2}} \right\} \quad (29)$$

[0110]

10 where

[0111]

$$x = \Delta V / \sqrt{K_3 I_0 / \beta}$$

[0112]

The above normalization enables the application not only to the first differential pair, made up of the transistors M1 and M2, but also to

15 the second differential pair, made up of the transistors M3 and M4.

[0113]

It is assumed that, with the first differential pair, comprised of the transistors M1 and M2,

[0114]

$$x_1 = \Delta V_1 / \sqrt{I_0 / (K_2 \beta)}$$

[0115]

and, with the second differential pair, made up of the transistors M3 and M4,

5 [0116]

$$x_2 = \Delta V_2 / \sqrt{K_3 I_0 / \beta}$$

[0117]

Since the drain current ID2 of the transistor M2 is K-tupled to  
10 flow through the transistor M4, with the current ratio in the second  
current mirror circuit being K3, the normalized input voltages are equal  
to each other, so that  $x_1 = x_2$ . Consequently,

[0118]

$$\Delta V_2 = \Delta V_1 \frac{\sqrt{K_3 I_0 / \beta}}{\sqrt{I_0 / (K_2 \beta)}} = \sqrt{K_2 K_3} \Delta V_1 \quad (30)$$

[0119]

15 with the multiplication coefficient being

[0120]

$$\sqrt{K_2 K_3}$$

[0121]

Since

$$20 \quad \Delta V_1 = \Delta V = \Delta V_{BE} = V_T \ln(K_1) \quad (31)$$

[0122]

$$V_{REF} = V_{BE2} + \Delta V_2 = V_{BE2} + \sqrt{K_2 K_3} V_T \ln(K_1) \quad (32)$$

[0123]

5           The thermal voltage  $V_T$  has a positive temperature characteristic of  $0.0853\text{mV}/^\circ\text{C}$ . If, assuming that the transistor  $Q_2$  is being driven with the constant current  $I_0$ , exhibiting a low temperature characteristic, the temperature characteristic of the base-to-emitter voltage  $V_{BE}$  of the transistor  $Q_2$  is  $-2.0\text{ mV}/^\circ\text{C}$ , and if it is desired for the output reference  
10   voltage  $V_{REF}$  not to display a temperature characteristic, it is sufficient if the temperature characteristic are cancelled out by a voltage exhibiting a positive temperature characteristic and a voltage exhibiting a negative temperature characteristic.

[0124]

15           That is, the value of  $\text{Sqrt}(K_2 \times K_3) \times \ln(K_1)$  is 23.447, where the function  $\text{Sqrt}()$  denotes  $\sqrt{\quad}$ .

          The value of  $\text{Sqrt}(K_2 \times K_3) \times V_T \ln(K_1)$  is  $0.60\text{ V}$  at ambient temperature.

          If  $V_{BE2}$  is  $0.7\text{V}$ ,

20            $\{V_{BE2} + \text{Sqrt}(K_2 K_3) \times V_T \ln(K_1)\} = 1.3\text{V}$ .

Specifically,

$K_1 = 10$ ,

$K_2 = 8$ , and



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In Fig.8, transistors M1 to M7 constitute a voltage follower type operational amplifier with a resistance for compensation RC and a capacity for compensation CC. The W/L ratio of input differential transistors M1 and M2 is set to 1:K2, and the W/L ratio of active load transistors M3, and M4, operating as loads, is set to K3:1, so that input offset is produced. The transistors M1 and M2, having sources connected in common to a drain of the constant current source transistor M5, form a differential pair. The transistor M3, connected to the drain of the current source transistor M1, and having its source grounded, and the transistor M4, having its drain connected to the drain of the transistor M2, having its source grounded and having its gate connected to the gate of the transistor M3, form a current mirror circuit operating as a load for the differential pair. The drain of the transistor M2, forming an output of the differential pair, has its source grounded, while having its drain connected to the gate of the transistor M5, the drain of which is connected to a drain of the constant current source transistor M7. An output voltage VREF is taken at a drain of the transistor M5, operating as an output terminal. The output terminal is connected to the gate of the transistor M2 that operates as an inverting input terminal of

SUB A13 7 the differential pair. A resistor RC for phase compensation and a capacity CC are connected across the drain and the gate of the transistor M5. A base-to-emitter voltage VBE of the transistor Q1 is input to a non-inverting input terminal of the differential pair.

5 [0127]

Since a correct resistance value is not required of the resistor RC for phase compensation, the resistor is usually replaced by a P-channel MOS transistor and an N-channel MOS transistor.

[0128]

10 The drain currents ID1 and ID2 of the respective transistors M1 and M2 are given by:

$$I_{D1} = \beta (V_{GS1} - V_{TH})^2 \quad (33)$$

$$I_{D2} = K3 \beta (V_{GS2} - V_{TH})^2 \quad (34)$$

The following relationship holds:

$$15 \quad I_{D1} + I_{D2} = I_0 \quad (35)$$

[0129]

Also, we can postulate:

$$V_{OS} = V_{GS1} - V_{GS2} \quad (36)$$

[0130]

20 Moreover, from the conditions of the active load transistors M3 and M4, we have

$$K3 I_{D1} = I_{D2} \quad (37)$$

Solving the equations (35) and (37),

$$I_{D1} = I_0 K3 / (K3 + 1) \quad (38)$$

$$25 \quad I_{D2} = I_0 / (K3 + 1) \quad (39).$$

[0131]

Thus, solving the above equation, we obtain:

[0132]

$$V_{os} = \sqrt{\frac{I_0}{\beta}} \sqrt{\frac{K_3}{K_3 + 1}} \left(1 - \frac{1}{\sqrt{K_2 K_3}}\right) \quad (40)$$

[0133]

5 Since the respective terms containing  $K_j$  are constants not dependent on the temperature, the temperature characteristic of the term of  $\text{Sqrt}(I_0/\beta)$  are at issue. Since mobility  $\mu$  exhibits temperature dependent characteristic in the case of the MOS transistor, the temperature dependency of the transconductance parameter  $\beta$  may be given by the following equation:

[0134]

$$\beta = \beta_0 \left(\frac{T}{T_0}\right)^{-\frac{3}{2}} \quad (41)$$

[0135]

where  $\beta_0$  is the value of  $\beta$  at an ambient temperature (300K). Of the temperature characteristic of the term  $\text{Sqrt}(I_0/\beta)$ , the temperature characteristic of the term  $\beta$  have become apparent. Next, the temperature characteristic of the constant current  $I_0$  must be defined.

[0136]

The generally used MOS reference current circuit may be implemented by self-biasing non-linear current mirror circuits, such as Nagata current mirror circuit, Widlar current mirror circuit, or an reverse Widlar current mirror circuit, as shown in Fig.8.

[0137]

Fig.8 shows a MOS reference current circuit, which is made up of a self-biased Nagata current mirror circuit.

[0138]

5 A circuit comprised of a transistor M10, having a source grounded, a drain connected to one end of a resistor R1 and having a gate connected to the opposite end of the resistor R1, a transistor M11, having a source grounded and having a gate connected to the drain of the transistor M10, and the resistor R1, makes up a Nagata current mirror circuit. Here, with the transistors M13 and M12 forming a current source, the transistors M10 and M11 and the resistor R1 form a self-biased Nagata current mirror circuit.

[0139]

Here, the transistor M10 is assumed to be a unit transistor, and the ratio of the gate width W to gate length L, or (W/L), of the transistor M11, is assumed to be K1 times that of the unit transistor, where  $K1 > 1$ .

[0140]

In the MOS Nagata current mirror circuit, shown in Fig.8, the device is assumed to exhibit satisfactory matching, the channel length modulation and the substrate effect are neglected, and the relationship between the drain current and the gate-to-source voltage of the MOS transistor is assumed to follow the square rule. Then, the drain current  $I_{D1}$  of the MOS transistor M10 is given by:

$$I_{D1} = \beta (V_{GS10} - V_{TH})^2 \quad (42).$$

25 [0141]

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The drain current  $I_{D2}$  of the MOS transistor M11 is given by:

$$I_{D2} = K1 \beta (V_{GS11} - V_{TH})^2 \quad (43)$$

There is also the following relationship:

$$V_{GS10} = V_{GS11} + R1 I_{D10} \quad (44).$$

5 [0142]

Solving the equation (44) from the equation (42), the relationship between the input and output currents of the MOS Nagata current mirror circuit is given by:

[0143]

$$I_{D11} = K1 \beta R1^2 I_{D10} \left( \sqrt{I_{D10}} - \frac{1}{R1 \sqrt{\beta}} \right)^2 \quad (45)$$

10 [0144]

[0145]

The MOS Nagata current mirror circuit features a region where the output current (mirror current) is monotonously increased against the input current (reference current), a peak point and a region where the output current (mirror current) is monotonously decreased against the input current (reference current).

15

[0146]

On differentiating  $I_{D11}$  with respect to  $I_{D10}$ , the  $I_{D11}$  reaches a peak point at

$$20 \quad I_{D11} = K1 \times I_{D10}/4 \quad \text{for } I_{D10} = 1/(4R1^2 \beta).$$

[0147]

Thus, if  $K1=4$ ,

$$I_{D11} = I_{D10}.$$

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[0148]

It is noted that transistors M15 and M14 make up a current mirror circuit, while the transistors M15 and M14 drive M10 and M11, respectively. Thus, the transistors M15 and M14 make up a MOS self-biased Nagata reference current circuit, with

$$I_{D10} = I_{D11} \quad (46)$$

Thus,

$$\Delta V_{GS} = V_{GS10} - V_{GS11} = R_1 I_{D10} \quad (47)$$

Solving the equation (39) from the equation (37),

[0149]

$$I_{D10} = I_{D11} = \frac{1}{R_1^2 \beta} \left(1 - \frac{1}{\sqrt{K_1}}\right)^2 \quad (48)$$

[0150]

where  $K_1$  is a constant not having a temperature characteristic.

Since mobility  $\mu$  exhibits a temperature characteristic in the case of the MOS transistor, the temperature dependency of the transconductance parameter  $\beta$  may be represented by the following equation:

[0151]

$$\beta = \beta_0 \left(\frac{T}{T_0}\right)^{-\frac{3}{2}} \quad (49)$$

[0152]

$\beta_0$  is the value of  $\beta_0$  at an ambient temperature (300K).

Therefore,

[0153]

$$\frac{1}{\beta} = \frac{1}{\beta_0} \left( \frac{T}{T_0} \right)^{\frac{3}{2}} \quad (50)$$

[0154]

The temperature coefficient of  $1/\beta$  is 5000 ppm/°C at ambient temperature. This value is approximately 1.5 times 3333 ppm/°C, which is the value of the temperature coefficient of the thermal voltage  $V_T$  of the bipolar transistor.

[0155]

Also, the transistor M12 forms a current mirror circuit with the transistor M13, so that

$$I_{D12} = I_{D13} \quad (51).$$

[0156]

That is, the output current  $I_0$  of the CMOS reference voltage circuit may be found by:

[0157]

$$I_0 = I_{D10} = I_{D11} = \frac{1}{R_1^2 \beta_0} \left( \frac{T}{T_0} \right)^{\frac{3}{2}} \left( 1 - \frac{1}{\sqrt{K_1}} \right)^2 \quad (52)$$

[0158]

where  $K_1$  is a constant not having a temperature characteristic. The temperature characteristic of  $1/\beta$  is approximately proportional to temperature, as mentioned above, and is 5000 ppm/°C at ambient temperature. This value is approximately 1.5 times 3333 ppm/°C, which is the value of the temperature characteristic of the thermal voltage  $V_T$  of the bipolar transistor.

[0159]

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Thus, it may be seen that, if a temperature coefficient of the resistor R1 is less than or equal to 5000 ppm/°C and linear as regards the temperature, the drain current ID10 exhibits a positive temperature coefficient, with the output current I0 of the reference current circuit, output by the current mirror circuit, being proportional to temperature, so that the circuit operates as a PTAT current source circuit.

[0160]

Although a startup circuit is needed for starting-up the self-bias circuit, this is omitted for simplicity in the previous explanation on the operation. For a simplified startup circuit, references may be made to the publication such as JP PATENT KOKAI JP-A-8-314561(Patent 2800720), by the same inventor as the present application .

[0161]

The output current of the CMOS reference current circuit is as shown in the equation (52), while its temperature characteristic has now become apparent. Consequently, by substituting the equation (52) into the equation (40), we obtain:

[0162]

$$\begin{aligned}
 V_{os} &= \frac{1}{R_1 \beta} \left(1 - \frac{1}{\sqrt{K_1}}\right) \sqrt{\frac{K_3}{K_3 + 1}} \left(1 - \frac{1}{\sqrt{K_2 K_3}}\right) \\
 &= \frac{1}{R_1 \beta_0} \left(\frac{T}{T_0}\right)^{\frac{3}{2}} \left(1 - \frac{1}{\sqrt{K_1}}\right) \sqrt{\frac{K_3}{K_3 + 1}} \left(1 - \frac{1}{\sqrt{K_2 K_3}}\right) \quad (53)
 \end{aligned}$$

[0163]

where the respective terms containing Kj are constants not exhibiting a temperature characteristic and an offset voltage VOS is determined by



the resistor R1 determining the current value of the CMOS reference voltage circuit, the temperature characteristic of  $1/\beta$ , which is 5000 ppm/°C at ambient temperature. If the temperature characteristic of the resistor R1 is sufficiently smaller than 5000 ppm/°C, the offset voltage exhibits a temperature characteristic of 5000 ppm/°C at ambient temperature. This value is approximately 1.5 times 3333 ppm/°C, which is the value of the temperature characteristic of the thermal voltage  $V_T$  of the bipolar transistor. Consequently, a reference voltage lower than the output voltage of the reference voltage circuit shown in Fig.7 is obtained as now explained.

[0164]

In Fig.8, the output voltage  $V_{REF}$  of the reference voltage circuit is given by:

$$V_{REF} = V_{BE1} + V_{OS} \quad (54).$$

[0165]

The transistor Q1 is driven with a constant current having a temperature characteristic of approximately 5000 ppm/°C.

[0166]

Therefore, if it is assumed that the temperature characteristic for the  $V_{BE}$  of the bipolar transistor is slightly less stringent than -1.9 mV/°C as explained with reference to Fig.7, and is slightly smaller than -1.9 mV/°C, for example, in the vicinity of -1.85 mV/°C, the temperature characteristic of the output voltage  $V_{REF}$  of the reference voltage circuit is given by:

[0167]

$$\frac{1}{R_1 \beta_0} \left(1 - \frac{1}{\sqrt{K_1}}\right) \sqrt{\frac{K_3}{K_3 + 1}} \left(1 - \frac{1}{\sqrt{K_2 K_3}}\right) = 0.37V \quad (55)$$

in case the temperature characteristic of VBE1 of -1.85 mV/°C and that of VOS of 500 ppm/°C cancel each other.

[0168]

If, in this case, VBE1 = 0.7 V, the output voltage VREF of the reference voltage circuit shown in Fig.8 is given by:

$$V_{REF} = 1.07V \quad (56).$$

[0169]

Since the circuit of the present embodiment takes the configuration of a voltage follower type operational amplifier, it is possible to subtract the offset voltage. In this case, the connection of various circuit components may be kept unchanged as shown in Fig.8 and only the gate W/L ratio of the transistors M1 and M2 and the gate W/L ratio of the transistors M3 and M4 are changed to K2:1 and to 1:K3, respectively. The output voltage VREF of the reference voltage circuit in this case is given by:

$$V_{REF} = V_{BE1} - V_{OS} \quad (57).$$

[0170]

Thus, if the offset voltage shown by the equation (57) is subtracted, the output voltage VREF of the reference voltage circuit for VBE1 = 0.7 V is given by:

$$V_{REF} = 0.33V \quad (58)$$

it being noted that the output voltage VREF of the reference voltage circuit exhibits a negative temperature characteristic of -3.7 mV/°C.

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[0171]

Fig.9 shows a modification of the embodiment shown in Fig.8. The drain and the gate of the transistor M2 of the differential pair are connected together and the output VREF is taken out from the drain. In Fig.9, the output voltage VREF of the reference voltage circuit is given by:

$$V_{REF} = V_{BE} + V_{OS},$$

as in equation (54), where  $V_{OS}$  is given by the equation (53).

That is, a reference voltage not dependent upon temperature is output, as mentioned above. Although this modification lacks in capability of feeding a current from the reference voltage output terminal, it is effective as a voltage source for supplying the reference voltage.

[0172]

In the above-described embodiments, the diode-connected bipolar transistors Q1 and Q2 may well be replaced by diodes. A Bi-CMOS circuit in which the bipolar transistor and the MOS transistors are constructed on one and the same substrate may also construct it. Although the present invention has been explained based on the above-described embodiments, the present invention is not limited to the particular structure shown therein and may, of course, be modified or corrected by those skilled in the art within the scope of the invention as defined by the claims.

[0173]

The meritorious effects of the present invention are summarized

as follows.

As described above, the present invention gives the following effects:

[0174]

5 The first effect of the present invention is that a reference voltage circuit having an output voltage of 1.2 V and not exhibiting a temperature characteristic may readily be implemented by the CMOS process.

[0175]

10 The reason is that, in the reference voltage circuit of the present invention, the circuitry is constructed measly by active devices, without employing resistors as in the conventional circuitry shown in Fig.10.

[0176]

15 The second effect of the present invention is that the reference voltage circuit not exhibiting a temperature dependent characteristic and having an output voltage lower than 1.2 V may be implemented by the CMOS process.

[0177]

20 The reason is that, in the reference voltage circuit of the present invention, the positive temperature characteristic of the bipolar transistor is canceled out by the negative temperature characteristic of the bipolar transistor of  $-1.9 \text{ mV}^\circ\text{C}$  by exploiting a temperature characteristic of  $5000 \text{ ppm}^\circ\text{C}$  derived from the term of  $1/\beta$ .

25 It should be noted that other objects, features and aspects of the present invention will become apparent in the entire disclosure and that

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modifications may be done without departing the gist and scope of the present invention as disclosed herein and claimed as appended herewith.

Also it should be noted that any combination of the disclosed and/or claimed elements, matters and/or items might fall under the  
5 modifications aforementioned.

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